ELECTRICAL ENGINEERING RESEARCH LABORATORY
ENGINEERING EXPERIMENT STATION
UNIVERSITY OF ILLINOIS
URBANA, ILLINOIS

QUARTERLY PROGRESS REPORT

on

RESEARCH STUDIES ON PROBLEMS
RELATED TO ANTENNAS

Quarterly Report No. 1

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Report for the Period

1 September 1962

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CONTENTS

1. RADIATION FROM PERIODIC STRUCTURES

1.1 Purpose

The purpose of this investigation is to explain the observed properties of log-periodic antennas through experimental and analytical investigation of their uniform periodic counterparts.

1.2 Factual Data

1.2.1 Research Staff - Paul E. Mayes, P. G. Ingerson

1.2.2 Status

Additional near-field measurements have been made on the uniform periodic dipole array described in Quarterly Report No. 4, Contract No. AF33(657)-8460. The half-length of the dipoles has been adjusted from a maximum of 28 cm. to a minimum of 16 cm. As expected, the major result of this change in length, which varies the resonant frequency of the dipoles, is to shift the stopbands. Figure 1.1 shows passband portions of the Brillouin diagram for the dipole array with twisted feeder for several dipole lengths.

A new uniform dipole array was constructed with elements which could be rotated on the twin-boom feeder. With this arrangement it is possible to determine the effect of the twisted feeder on the Brillouin diagram of the structure. Figure 1.2 shows the Brillouin diagrams for the same dipole arrays using the twisted and straight feed lines. Since the difference between these two structures is primarily in the mutual impedance, a comparison of these two diagrams shows the influence of the twisted feeder on the coupling between elements.

Some data were also taken with the dipole halves rotated to be parallel so that they form open transmission line stubs rather than dipole antennas. Most points in the passband region of the Brillouin diagram for the stub-loaded line were found to be very close to the points on the dipole-loaded line with twisted feeder. A comparison of these two diagrams is shown in Figure 1.3.

During this quarter a paper entitled, "Near Field Measurements on Backfire Periodic Dipole Arrays," was presented at the Fall meeting of URSI (International Union of Radio Scientists) in Ottawa, Canada and at the 12th Annual Symposium on the USAF Antenna Research and Development Program at Allerton

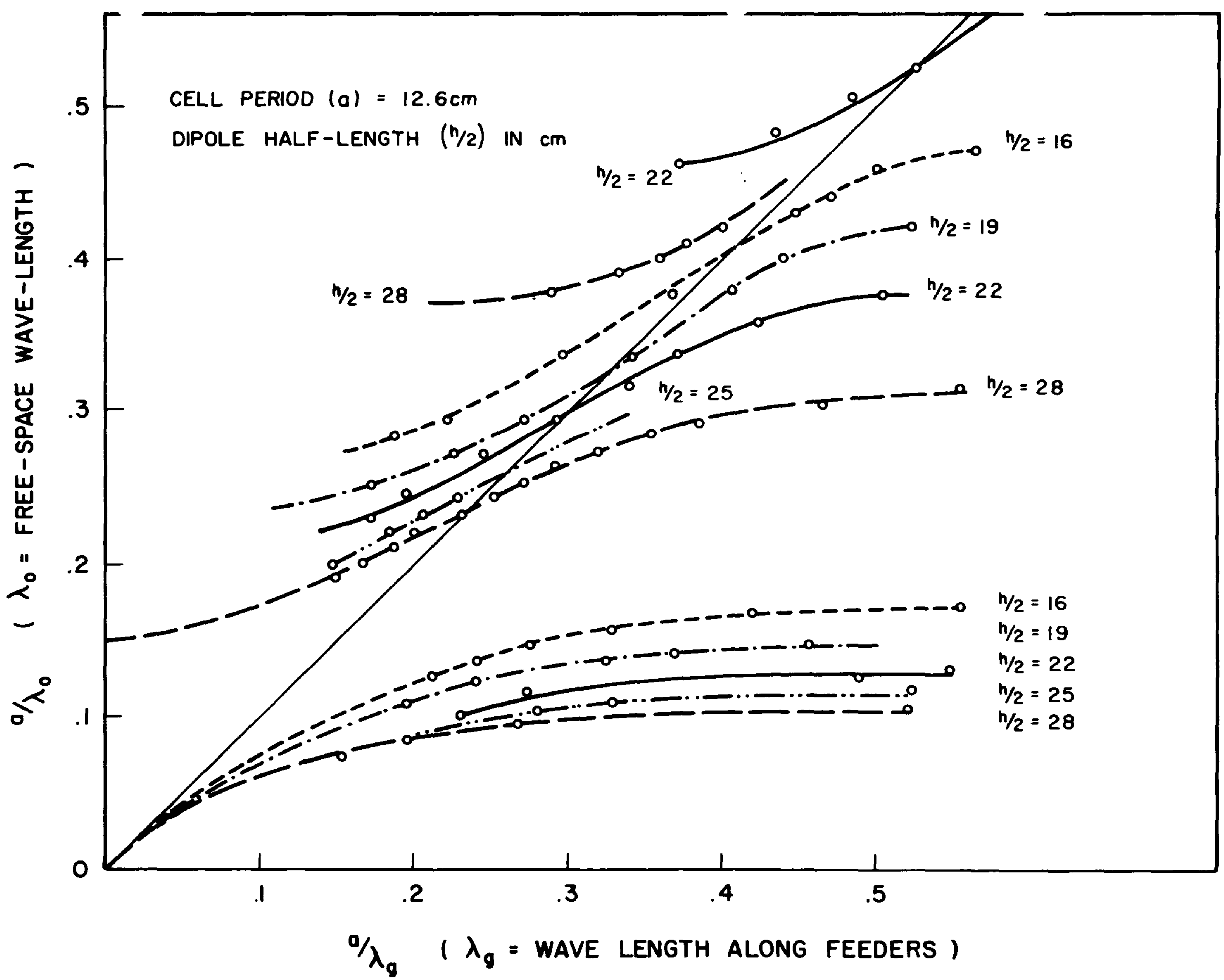


Figure 1.1. Brillouin Diagram Perturbation of Uniformly Dipole Array as Dipole Half-Length (h/2) is Varied

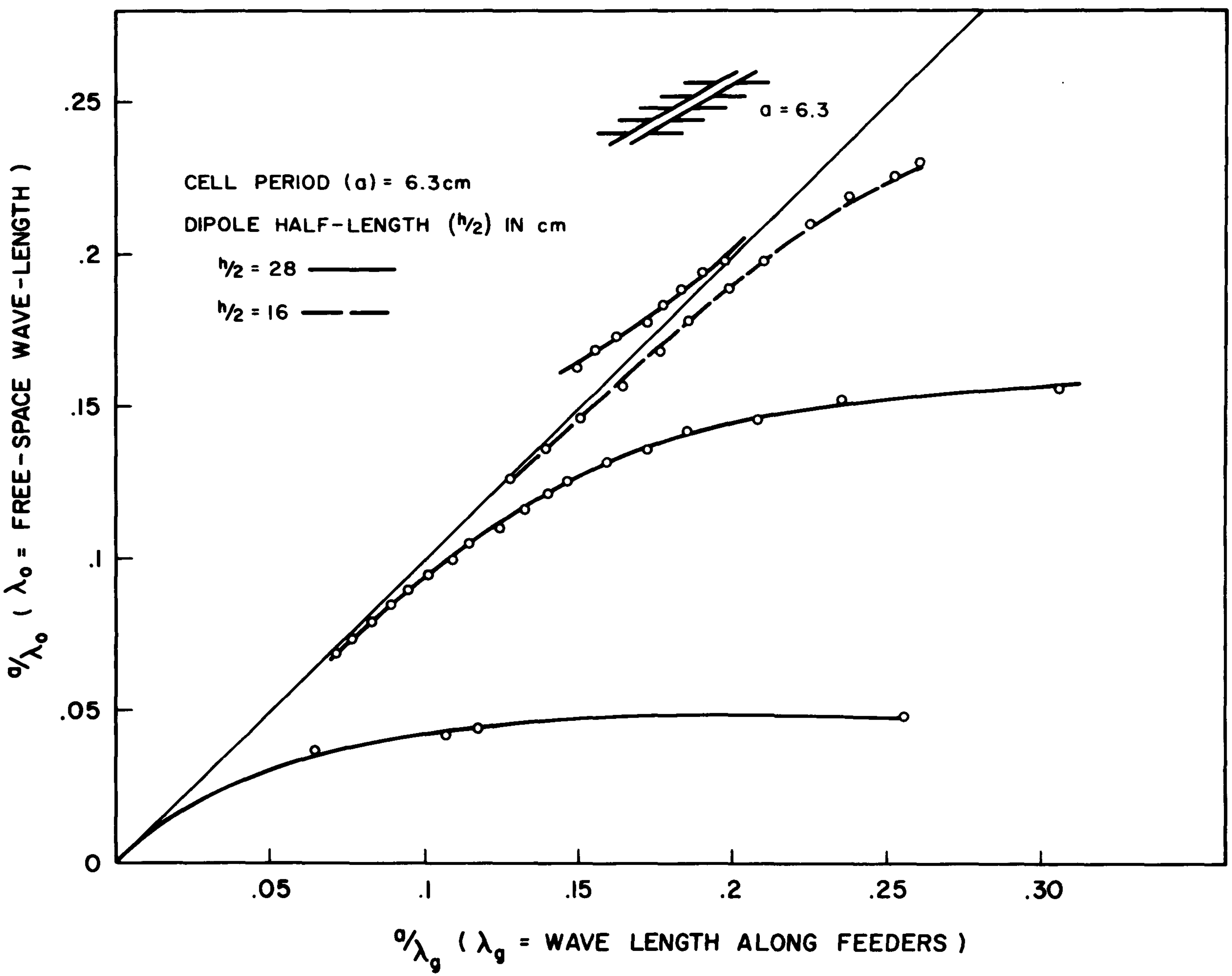


Figure 1.2. Brillouin Diagram for Periodic Mono-pole (Straight Feeders)

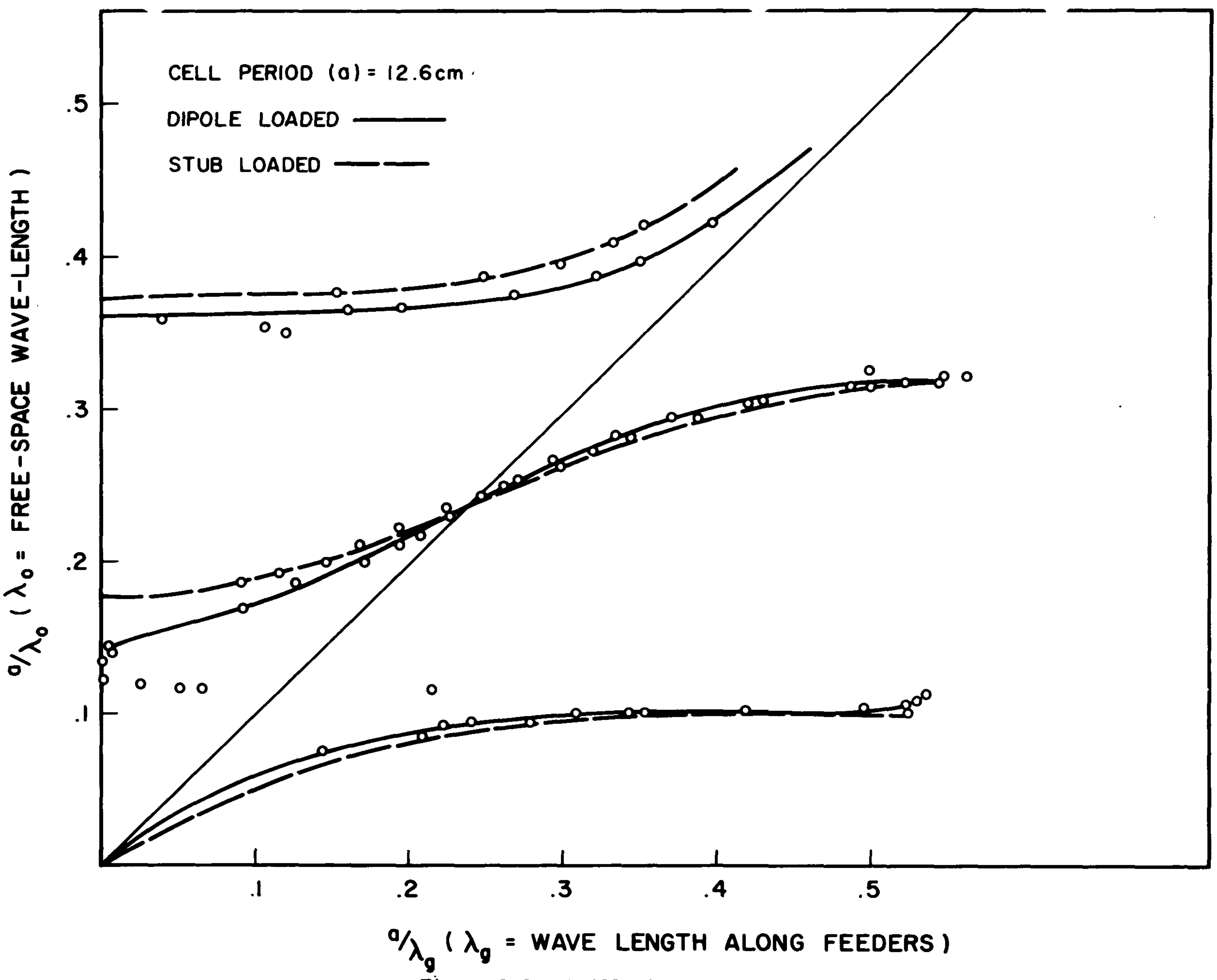


Figure 1.3. Brillouin Diagrams for Dipole and Stub Loaded Periodic Arrays

Park, Illinois. A technical report "Backward Wave Radiation from Periodic Structures and Application to Design of Frequency-Independent Antennas" was completed.

1.3 Plans for the Next Interval

Additional measurements will be made of the Brillouin diagram of stub-loaded lines. Correlation between measured radiation patterns and the near field measurements will be sought.

2. LOG-PERIODIC CAVITY-BACKED SLOT ANTENNAS

2.1 Purpose

This work will be directed toward development of a log-periodic array of cavity-backed slots which will produce vertically polarized radiation without having structure projecting above the ground plane.

2.2 Factual Data

2.2.1 Research Staff - P. E. Mayes and V. A. Mikenas

2.2.2 Status

This project was initiated during this quarter in an effort to apply some of the more recent concepts in log-periodic antenna development to the problem of flush-mounted antennas. As has been demonstrated in the case of the periodic dipole array, a line which is periodically-loaded with resonant elements will possess stopbands which cause rapid decay of the wave along the line. Introducing π radians additional phase shift between resonant radiating elements yields backfire radiation in the stop-band. The basic idea of this project is to achieve the same type of performance using cavity-backed slots as the resonant radiating elements rather than linear dipoles.

The type of cavity-backed slot under consideration is shown in Figure 2.1. The cavity is excited by a loop at its base and radiates through a rectangular slot in a ground plane at its top. The input impedance versus frequency of this resonant radiator resembles the input admittance versus frequency of linear dipoles. The cavity-slot is therefore the dual of the dipole. The dipoles represent shunt loading on the line of a log-periodic dipole array. The dual of the dipole array would therefore involve series loading. The Brillouin diagram of a periodic array of series-fed closed cavities should correspond to that of the line loaded with shunt stubs. Permitting radiation by cutting slots in the cavities will perturb the diagram due to losses and mutual coupling. The resulting diagram should correspond to the diagram of the dipole loaded line. Introducing π radians phase shift between adjacent cavities should produce backfire radiation in the cavity-backed slots as it does in the dipole array. A series-loop feed has been used successfully by Wahl and Mittra (see Section 9) to produce backfire radiation from a corrugated surface.

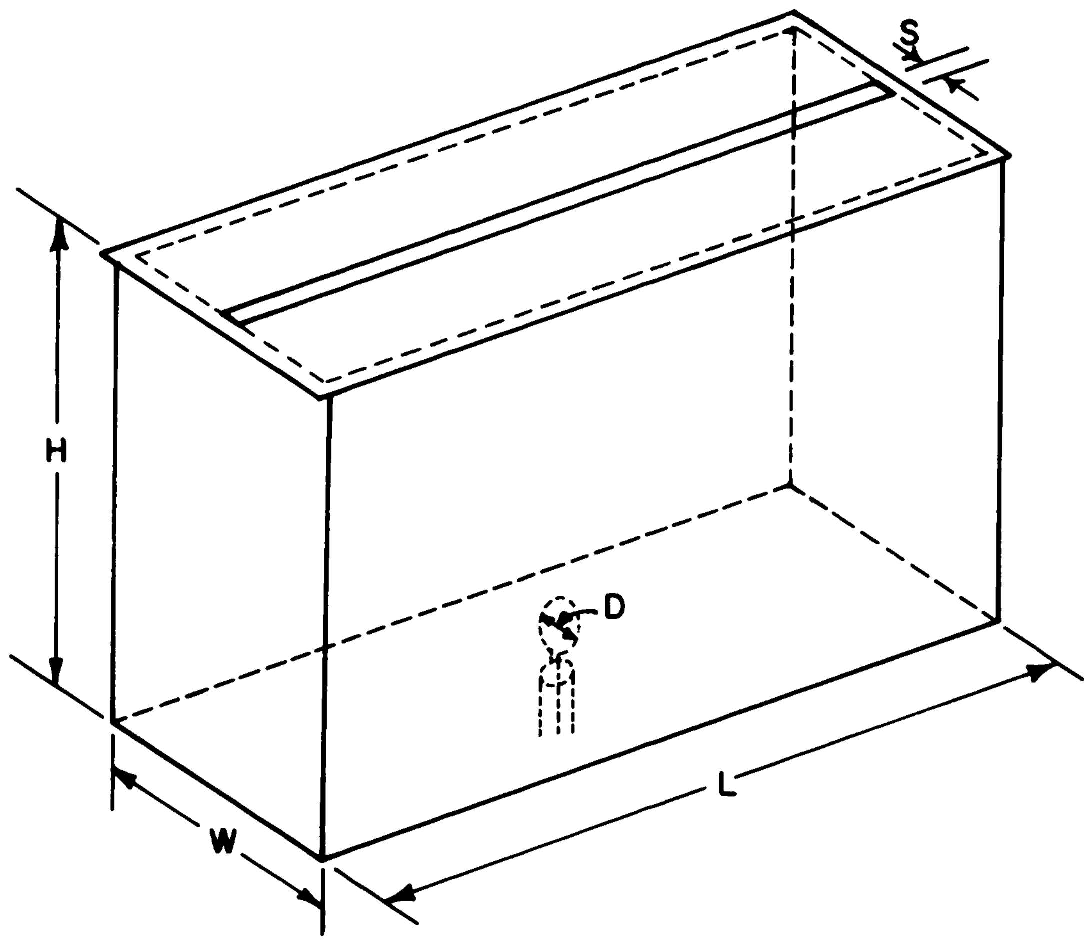


Figure 2-1

During this period measurements of input impedance have been made on the cavity-backed slots shown in Figure 2.1. A typical impedance versus frequency plot is shown on the Smith Chart in Figure 2.2, for a cavity with the following dimensions:

L = 30 cm
W = 8 cm
H = 15 cm
S = 1.25 cm
D = 2.75 cm

The slot width and loop radius have been varied to assess the effect of these parameters.

2.3 Plans for the Next Interval

A log-periodic array of cavity-backed slots will be constructed and tested. If time permits, a uniformly periodic array will also be constructed and near-field measurements made to determine the Brillouin diagram for the structure.

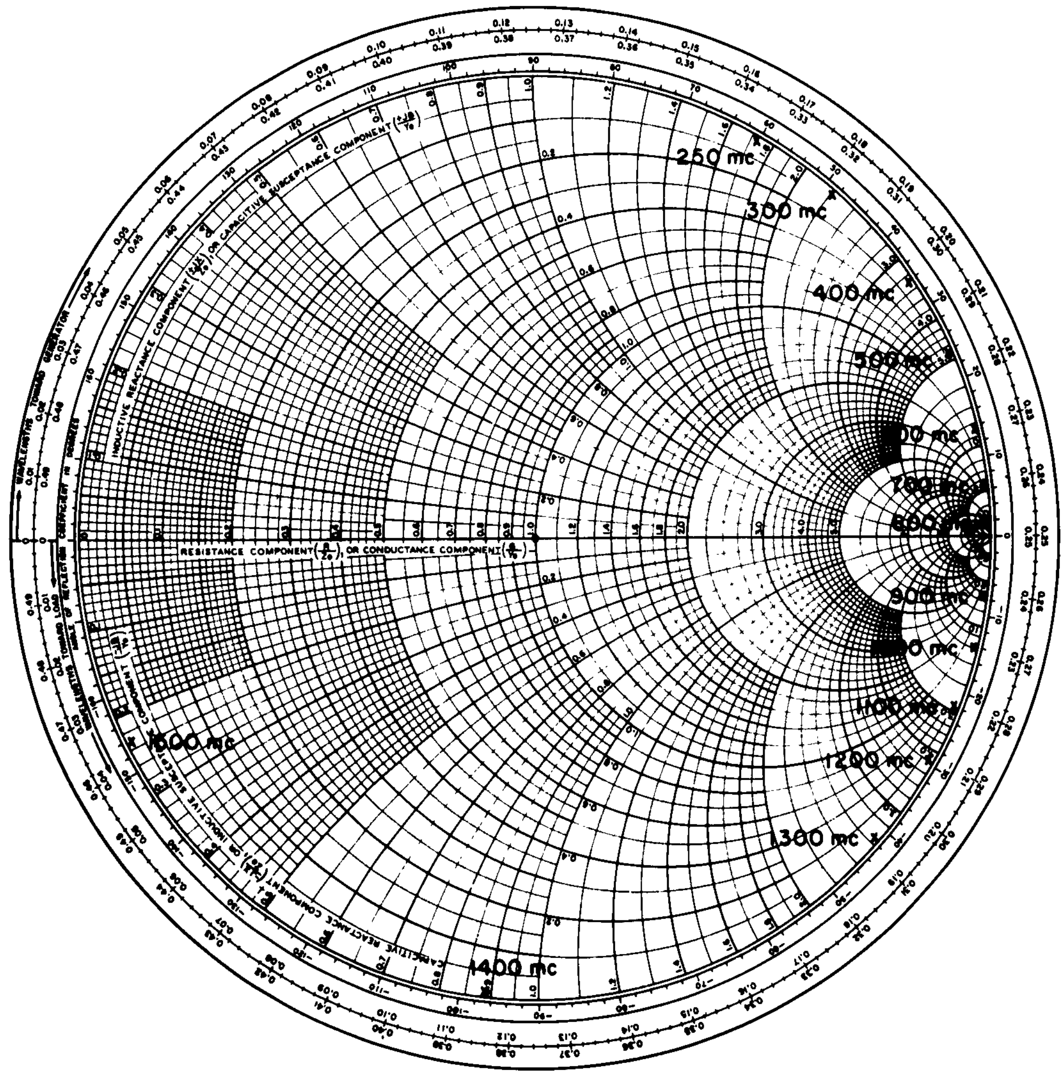

250 mc
300 mc
400 mc
500 mc
700 mc
900 mc
1100 mc
1200 mc
1300 mc
1400 mc
1600 mc
RESISTANCE COMPONENT

Figure 2-2

3. LOG-PERIODIC MAGNETIC-CURRENT ANTENNAS

3.1 Purpose

The purpose of this work is to investigate an LP array of magnetic dipole elements.

3.2 Factual Data

3.2.1 Research Staff - P. E. Mayes

3.2.2 Status

Measurements of input impedance with conducting cones inserted in the magnetic current antenna showed no appreciable effect due to the cone. Thus it was found that a cone with dimensions indicated in Figure 3.1 could be inserted without affecting appreciably either the radiation patterns or input impedance.

Since the input impedance is high, new measurements were made using the 300-ohm Lecher-wire system. The preliminary result from these measurements indicate that the VSWR with respect to 300 ohms is less than 3:1 over the frequency range from 455 to 610 Mc. The radiation patterns of the antenna are shown in Figure 3.2 and 3.3. The pattern bandwidth is seen to be somewhat greater than the impedance bandwidth.

3.3 Plans for the Next Interval

A more accurate model will be constructed to attempt to improve the impedance and extend the impedance bandwidth. Further data will be obtained on the influence of interior cones.

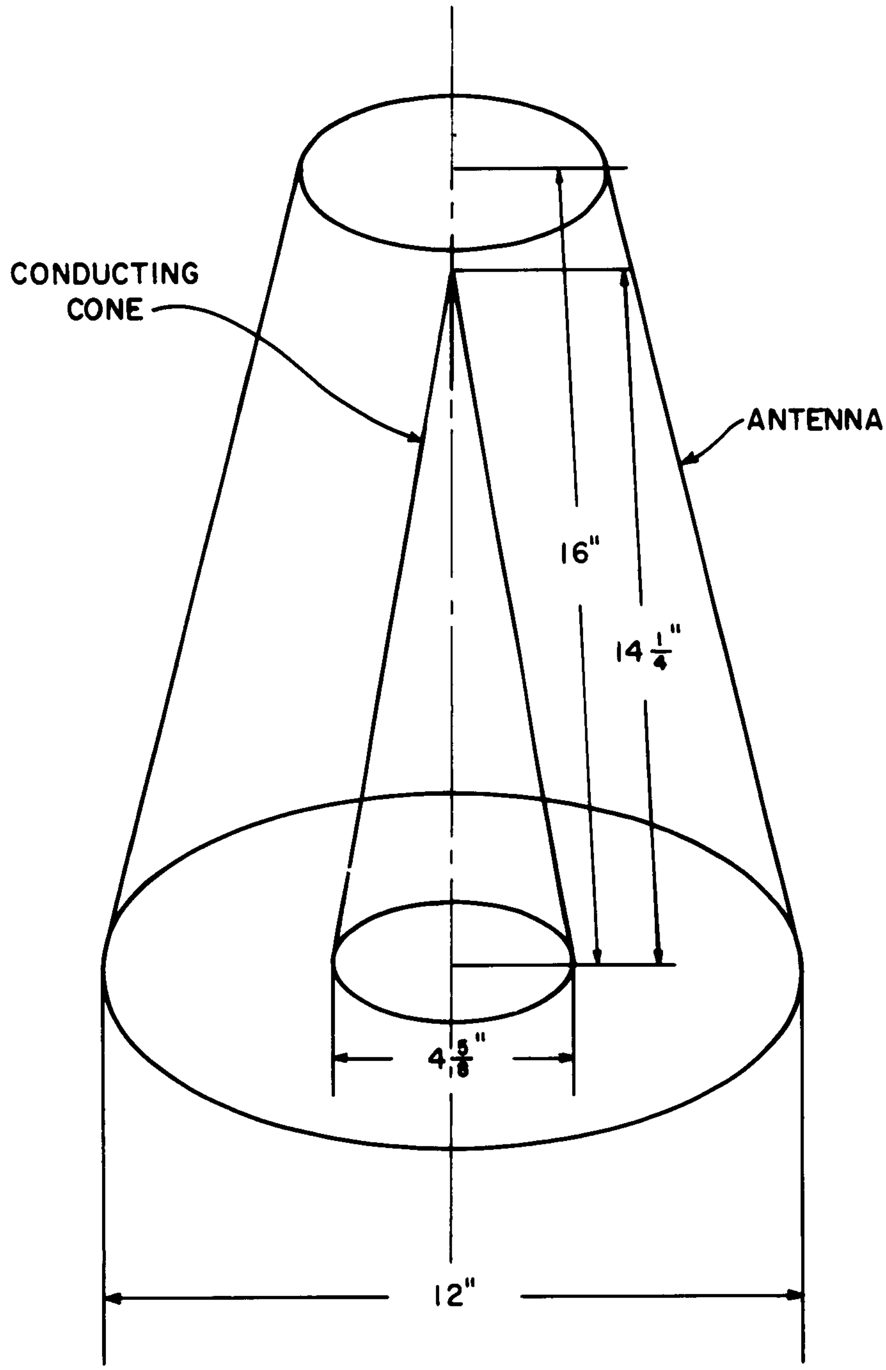


Figure 3.1. Dimensions of antenna and interior conducting cone.

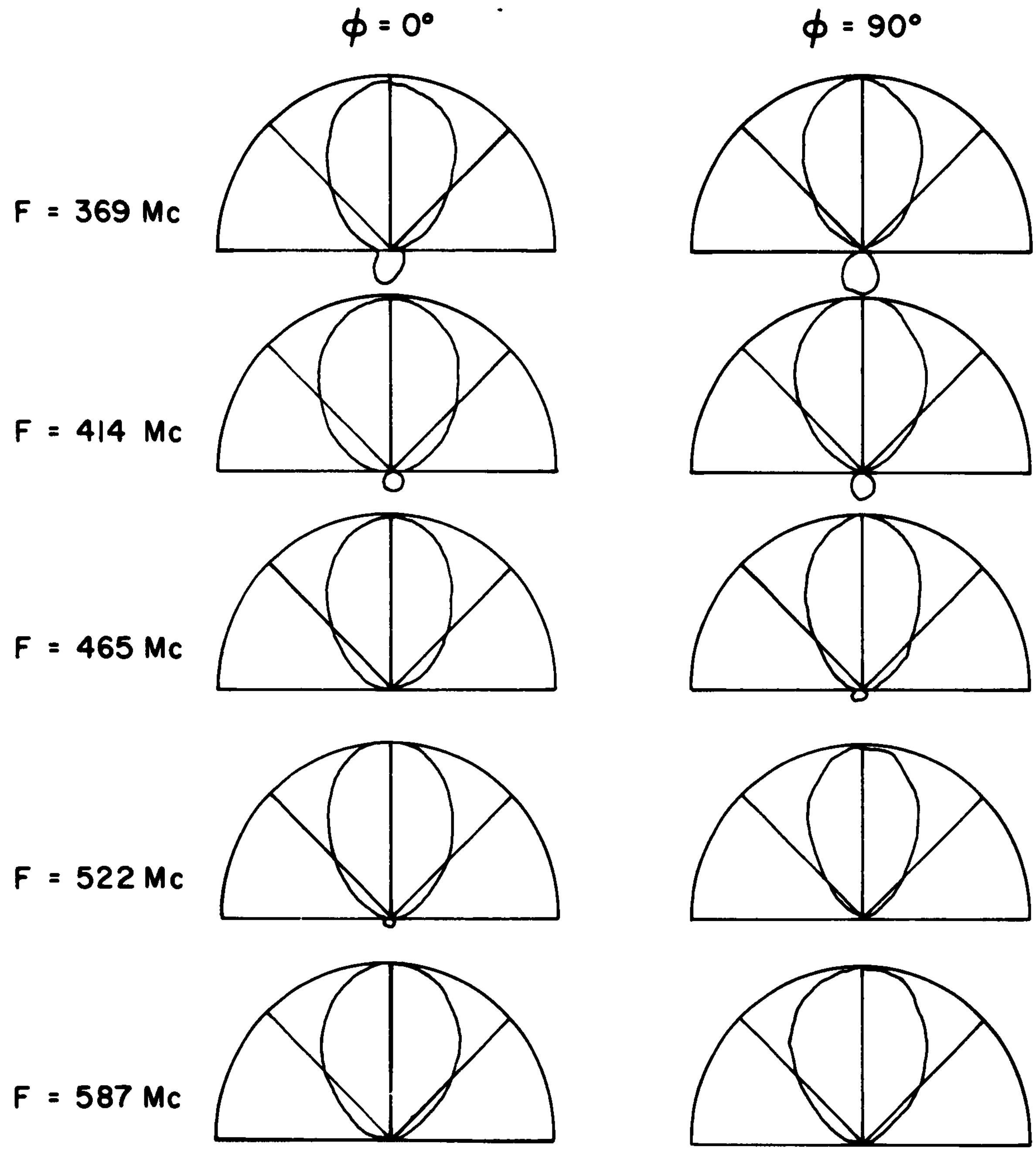


Figure 3.2. Radiation Patterns of LP Magnetic Current antenna (τ = 0.89, α = 0.943, cone angle 20°)

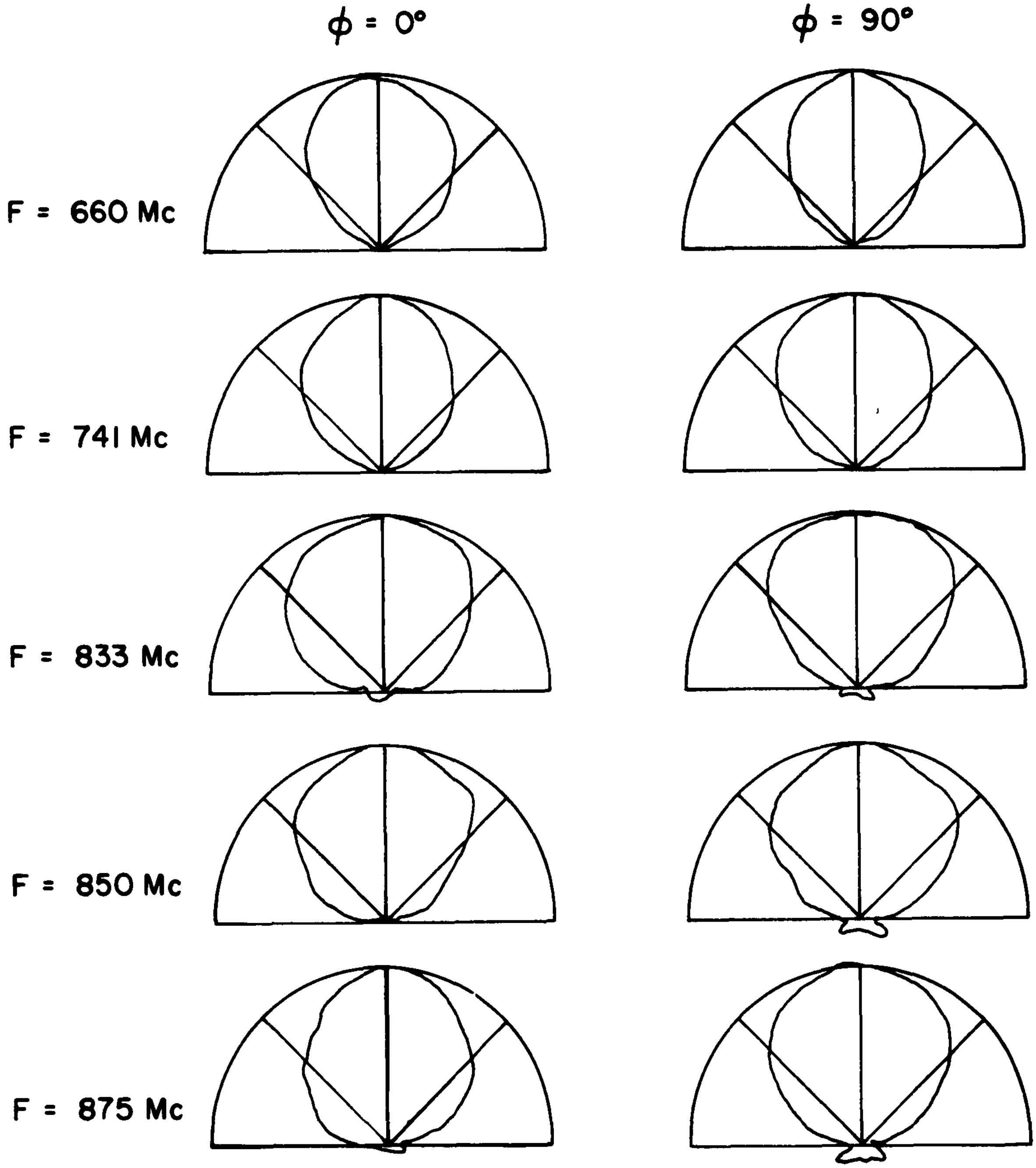


Figure 3.3. Radiation Patterns of LP Magnetic Current Antenna (τ = 0.89, α = 0.943, cone angle 20°)

4. LOG-SPIRAL ANTENNAS

4.1 Purpose

The objective of this investigation is to study the characteristics of present forms of the logarithmic spiral antennas and investigate new versions of the antenna.

4.2 Factual Data

4.2.1 Research Staff - J. D. Dyson

4.2.2 Status

As outlined in prior reports an intensive investigation is being made of the amplitude and phase characteristics of the near fields on the conical log-spiral antennas. The consideration of the conical logarithmic structure to be locally periodic with slowly varying parameters allows this experimental data to be analyzed in terms of prior work on the cylindrical monofilar and bifilar helix.

One cell of a cylindrical helical structure superimposed upon a corresponding cell of a wire version of the conical structure is shown in Figure 4.1. The pitch of the helix is given by

$$p = \ell \sin \psi = 2\pi a \tan \psi$$
$$= 2\pi a \cot \alpha$$

where ℓ is the length of one turn, ψ is the pitch angle and α its complement, the spiral angle, and "a" is the radius of the helix. The corresponding pitch distance of the conical spiral is given by

$$p' = \frac{2a}{\sin \theta_o} \sinh \left(\frac{\pi \sin \theta_o}{\tan \alpha}\right)$$

The pitch is the spacing between turns, or effectively the array element spacing of the cylindrical structure. The turn to turn phasing is determined, to the first order, by the ratio of the pitch to the turn length. To indicate the difference between the cylindrical and conical cells we can define a tapering factor which takes into account the pitch to turn length ratios of the two structures.

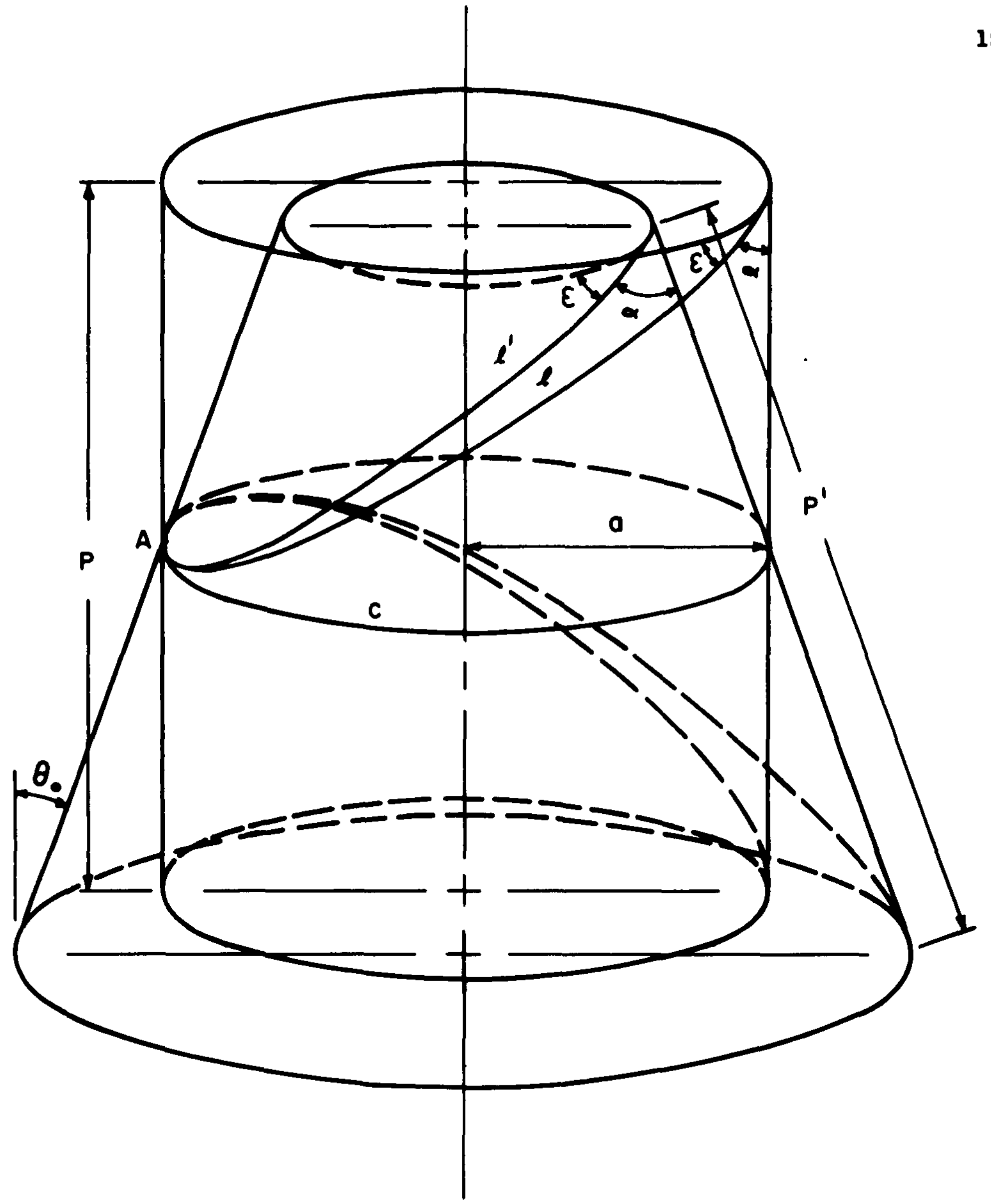


Figure 4.1. Cells of Conical Spiral and Cylindrical Helix Structures.

This tapering factor, indicating the difference or deviation of the pitch of the conical structure, p', from that of the cylindrical structure, p, for a given radius a is shown in Figure 4.2. The pitch of conical cells with combinations of parameters α and θ_o lying below the 2% curve will be within 2% of the pitch of the corresponding cylindrical cell. The antennas with well-formed beams and constant characteristics with frequency, i.e., $2\theta_o \leq 40^o$ and $\alpha \geq 72^o$ fall in this area.

The propagation constants along the surface of several conical antennas have been measured. Figure 4.3 is a fairly typical plot of the near field distribution obtained on one antenna when this field is measured with a small shielded loop. The two curves were recorded with the plane of the loop positioned parallel to the arms and moved along a line parallel to, and 2 mm and 30 mm away from, the surface of the arms. The horizontal coordinate axis of the graph is linear and is marked to show the approximate position of the arms. The region where the fields decay very rapidly scales up and down the structures as the frequency is varied.

4.3 Plans for the Next Interval

These measurements will be continued and extended. An attempt will be made to calculate the radiation patterns to be expected from the measured propagation constants and compare these with measured patterns. This should definitely establish those turns that are contributing significantly to the radiation pattern.

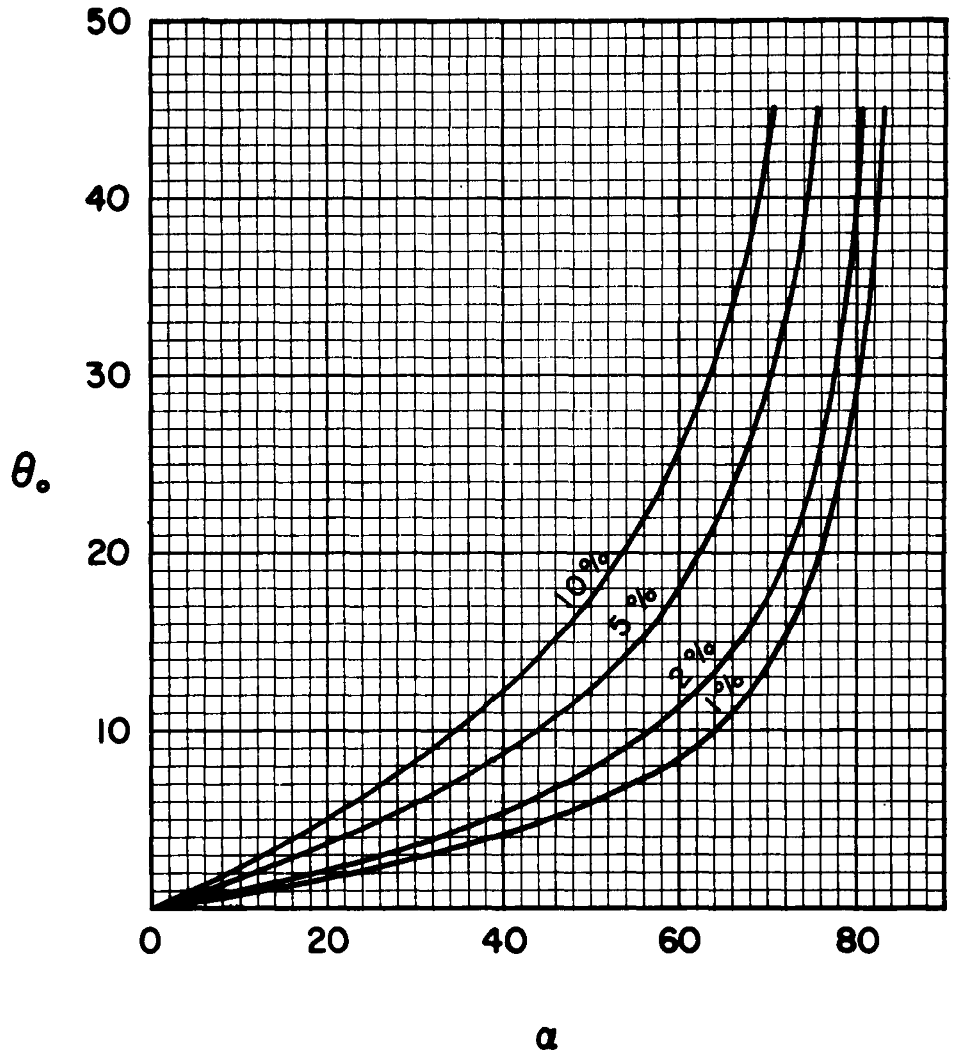


Figure 4.2. Tapering Factor for Conical Structures.

$$\frac{p' - p}{p} (100\%)$$

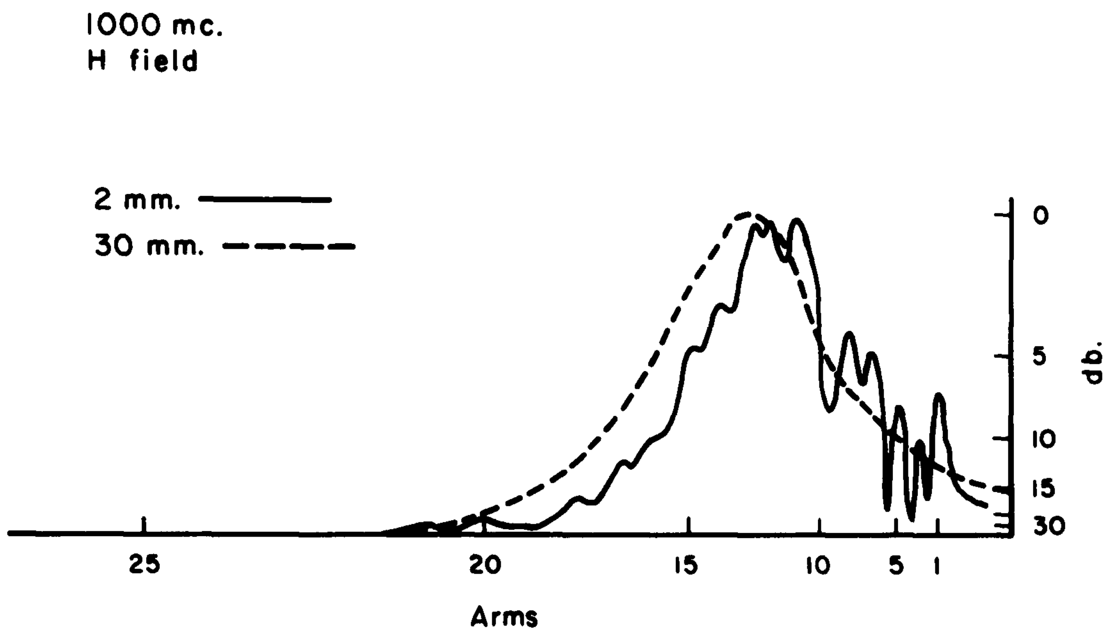


Figure 4.3. Near field along one conical log-spiral antenna measured with a shielded loop probe positioned 2 and 30 mm from surface of cone. $\alpha = 83^{\circ}$, $\theta_{o} = 10^{\circ}$.

5. LOG-PERIODIC ZIGZAG ANTENNAS

5.1 Purpose

This work is directed toward obtaining design information about a class of periodic leaky wave structures which are potentially high-gain, frequency-independent antennas.

5.2 Factual Data

5.2.1 Research Staff - P. E. Mayes

5.2.2 Status

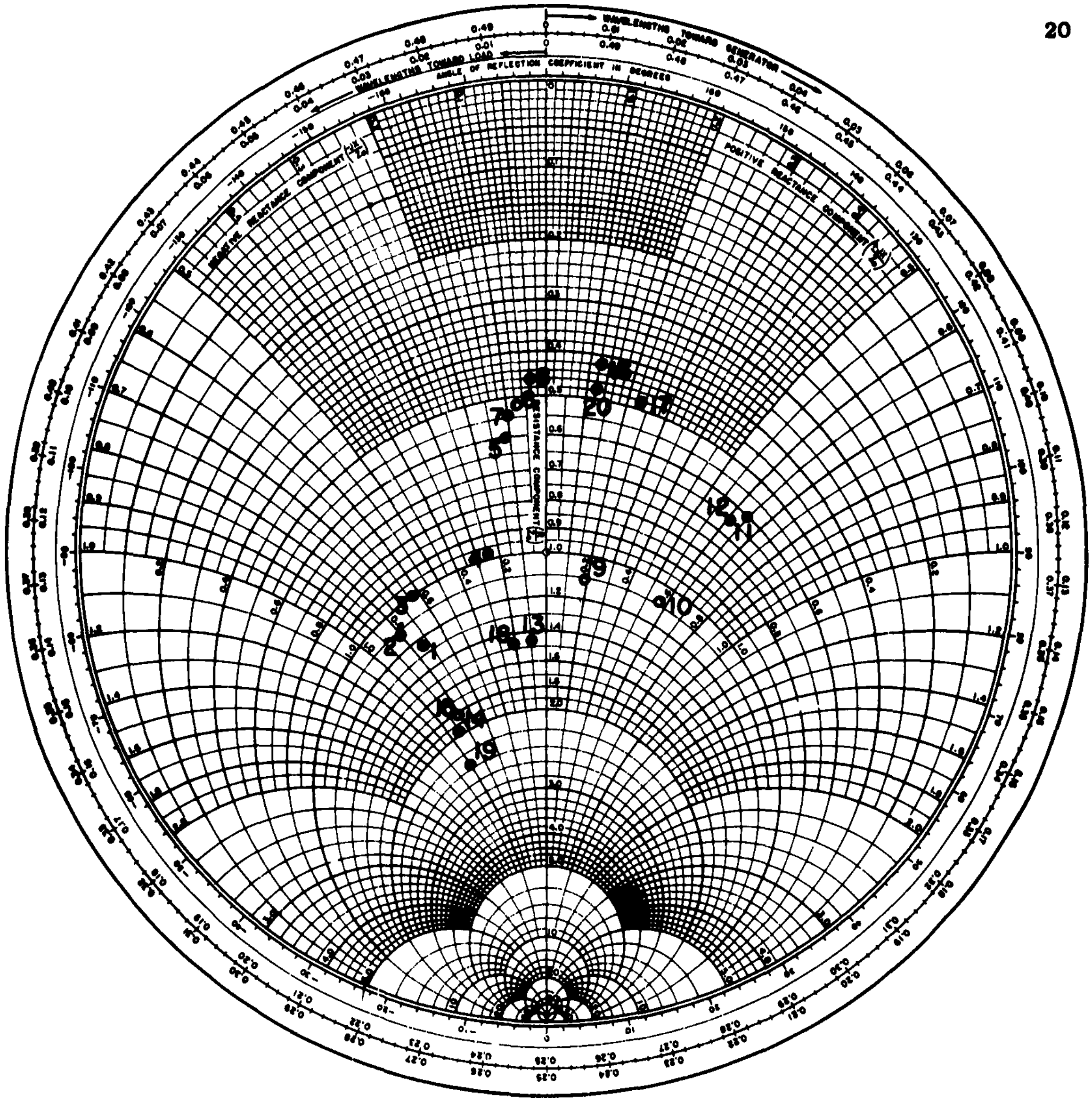
Additional impedance measurements have been made on balanced log-periodic zigzag antennas. Errors were found in the impedance plot reported in the preceding quarterly report (Quarterly Report No. 4, Contract No. AF33(657)-8460). The corrected impedance at the input terminals of the antenna (LPZZ-3, $\tau = 0.9$, $\alpha = 5^{\circ}$, $\psi = 10^{\circ}$) is plotted in Figure 5.1. The impedance is normalized to the characteristic impedance of the 300-ohm Lecher-wires used in the measurement. The mean resistance level is about 320 ohms and the maximum VSWR with respect to that level over a frequency band from 250 to 700 Mc is 2.7:1.

Preliminary measurements on other LP zigzag antennas indicate that the VSWR can be reduced by adjusting different parameters. There remains some question as to the effect of the wooden supports which have been used in constructing the antennas. The impedance measurements will be repeated with the antennas reconstructed with some of the wood removed in order to evaluate this effect.

Improved indoor facilities for impedance measurements are under construction. These new facilities will make it possible to continue the measurement program during the winter months ahead and should afford better accuracy.

5.3 Plans for the Next Interval

The impedance measurements will be extended to other parameters including uniform zigzags which correspond to pattern models previously tested (see Technical Report No. 60, AF33(657)-8460, "Backward Wave Radiation from Periodic Structures and Application to the Design of Frequency-Independent Antennas").



Points	Frequency	Points	Frequency
1	254.6 Mc	11	431.5 Mc
2	268.4	12	454.8
3	283.0	13	479.4
4	298.4	14	505.3
5	314.6	15	532.6
6	331.6	16	561.4
7	349.6	17	591.8
8	368.5	18	623.8
9	388.4	19	657.5
10	409.4	20	693.0

Figure 5.1. Measured input impedance of log-periodic zigzag ($\tau = 0.9$, $\alpha = 5^\circ$, $\psi = 10^\circ$). Normalization impedance = 300 ohms.

6. AN INTEGRATED ANTENNA AMPLIFIER

6.1 Purpose

This investigation will concern the feasibility of utilizing varactor diodes in a backfire antenna to produce amplification in the antenna itself and thereby improve signal-to-noise performance.

6.2 Factual Data

6.2.1 Research Staff - P. E. Mayes

6.2.2 Status

Varactor diodes have been obtained with cut off frequencies in the range from 8 to 75 Gc. The first integrated antenna-amplifier will be designed to operate near 300 Mc so that the signal-to-cut off frequency ratio f_s/f_c will be approximately 10^{-2}. Further information gained through the measurements reported in Section 1 may be used to make filter sections to separate signal, pump and idler frequencies.

6.3 Plans for the Next Interval

Initial testing will be done with a transmission line loaded periodically with both dipoles and diodes. The effect of adding diodes to the dipole-loaded line will be evaluated by the measurements.

7. COLLECTING SYSTEMS CONTAINING NONLINEAR AND/OR ACTIVE ELEMENTS

7.1 Purpose

The purpose of this investigation is to study the potentialities of collecting systems for electromagnetic waves in which active and/or non-linear elements are made part of the antenna.

7.2 Factual Data

7.2.1 Research Staff - R. H. MacPhie

7.2.2 Status

During the last quarter a solution to the problem of maximizing the signal-to-noise ratio at the terminals of a linear (i.e., one-dimensional) antenna has been obtained. Given a signal with a known direction of arrival and an arbitrary distribution of background noise sources, one wishes to maximize the average signal-to-noise power ratio at the terminals of the antenna. This can be expressed mathematically as

$$\frac{< v_s^2 >}{< v_N^2 >} = \text{MAX}$$

where v_s and v_N are respectively the signal and noise terminal voltages and the brackets $< \ldots\ldots >$ indicate a time average.

If one sets a limit on the amount of supergaining of the antenna, then the calculus of variations can be used to deduce an integral equation for the optimum aperture weighting function, $\bar{a}_o(x)$, of the antenna. It is given by

$$\int_{-L/2}^{L/2} N(x,y)\ \bar{a}_o(x)\ dx = e^{-ju_s y}$$

for

$$| y | \leq L/2$$

where $u_s = \beta \sin \theta_s$ is the direction of arrival of the desired signal, and $N(x,y)$ is the mutual coherence function of the narrow-band complex noise envelopes at the points x and y in the aperture of the antenna. The length of the aperture is L.

An exact solution to the integral equation is obtained by expanding the functions $N_o(x,y)$, $a_o(x)$, and $e^{ju_s y}$ as infinite series of functions orthonormal in the region occupied by the antenna. This leads to an infinite set of equations to be solved for the unknown coefficients of the expansion of $a_o(x)$. In practice this set will have to be truncated at some point and it is shown that by taking only $2(L/\lambda) + 1$ equations (L/λ is the antenna size in wavelengths), one obtains the optimum <u>practical</u> solution. By <u>practical</u> is meant that if the number of equations is increased beyond $2L/\lambda + 1$ an unstable supergain condition quickly develops which has little practical value.

<u>7.3 Plans for the Next Interval</u>

Work on this problem and on that of cross-correlation systems in general will be continued.

8. WAVE PROPAGATION ALONG HELICAL CONDUCTORS

8.1 Purpose

The purpose of this investigation is to calculate the complete k-β diagram (Brillouin diagram) for the helical structure and study the application of this diagram to helical radiators.

8.2 Factual Data

8.2.1 Research Staff - P. Klock, R. Mittra

8.2.2 Status of Work

During this interval two source problems were investigated. This first is an infinite helix with a source of electric field at the origin. For this problem the amplitudes of the free modes are calculated and the ratio of the magnitudes of the amplitudes for a specific helix is studied as a function of frequency.

The second problem is a finite helix fed in the center. Using the phase constants calculated from the determinantal equation a variational technique is used to calculate ratio of amplitudes. The ratio is studied for a specific helix as a function of frequency.

8.2.2.1 The First Problem

Let the component of electric field on the center line of the infinite helix be given by

$$E = \begin{cases} E_o & |\varphi| \leq \varphi_o/2 \\ 0 & |\varphi| > \varphi_o/2 \end{cases}$$

Then E may be written as a Fourier integral as

$$E_{\parallel} = \frac{E}{\pi} \int_{-\infty}^{\infty} \frac{\sin \beta \, \varphi_o/2}{\beta} \, e^{-j\beta\varphi} \, d\beta$$

since $E/\pi\beta \sin \beta\varphi_o/2$ is the Fourier transform of the assumed electric field. The current on the tape is

$$I(\varphi) = \int_{-\infty}^{\infty} I(\beta)\, e^{-j\beta\varphi}\, d\beta$$

Now E may be calculated from the current as

$$E = \frac{-j\omega\mu}{4\pi k^2 (a^2+\bar{p}^2)^{1/2}} \int_{-\infty}^{\infty} \left\{ \frac{\partial^2 I(\varphi')}{\partial \varphi'^2} + I(\varphi')\; k^2 [a^2 \cos(\varphi-\varphi') + \bar{p}^2] \right\} \frac{e^{-jkR}}{R}\, d\varphi'$$

Equating the two expressions for $E_{\parallel}$ and replacing the current and its derivative by their Fourier integral one obtains after an exchange of order of integration

$$0 = \int_{-\infty}^{\infty} d\beta \left\{ \frac{E}{\pi} \frac{\sin \frac{\beta\varphi_o}{2}}{\beta} e^{-j\beta\varphi} + j \frac{I(\beta)}{4\pi\omega\, \epsilon (a^2+\bar{p}^2)^{1/2}} \int_{-\infty}^{\infty} \left[\left\{ -\beta^2 + \right.\right.\right.$$

$$\left.\left.\left. + k^2 [\cos(\varphi-\varphi') + \bar{p}^2] \right\} \frac{e^{-jkR}}{R} e^{-j\beta\varphi'} \right] \left[d\varphi' \right] \right\}$$

Setting the integrand equal to zero one may solve for $I(\beta)$. The integral with respect to φ' is recognized as the electric field caused by a current of the form $e^{-j\beta\varphi}$. Consequently, the integral may be replaced by the electric field written as a series. The series when set equal to zero is the determinantal equation. The series expression is

$$E = \frac{j\, e^{-j\beta\frac{\delta}{2}}\, \bar{p}\, I_o\, v\, e^{-j\beta\varphi} \sin^2\psi}{2k\epsilon p} F(\beta)$$

where

$$F(\beta) = 2A_o \left[\beta^2 - \frac{k'^2}{\sin^2\psi} - k'^2 \operatorname{ctn}^2\psi \,\frac{A_{-1} + A_{+1} - 2A_o}{2A_o} \right]$$

$$k' = \frac{ka}{\operatorname{ctn}\psi}$$

$$A_j = \sum_{n=-\infty}^{\infty} I_{n+j}\left(\tau_n \frac{a}{\bar{p}}\right) K_{n+j}\left(\tau_n \frac{a}{\bar{p}}\right) D_n$$

$$D_n = \frac{\sin \beta_n \frac{\delta}{2\bar{p}}}{\beta_n \frac{\delta}{2\bar{p}}}, \qquad \beta_n = \beta + n, \qquad f = \frac{p}{2\pi}$$

$$\tau_n = (\beta_n{}^2 - k'^2)^{1/2}$$

$I(\varphi)$ is then found from $I(\beta)$ by

$$I(\varphi) = \int_{-\infty}^{\infty} I(\beta)\, e^{-j\beta\varphi}\, d\beta$$

where

$$I(\beta) = \frac{-j \frac{E\varphi_o}{\pi} \frac{k' \epsilon p}{v \sin^2 \psi} \frac{\sin \frac{\beta\varphi_o}{2}}{\frac{\beta\varphi_o}{2}} e^{-j \beta \frac{\delta}{2\bar{p}}}}{F(\beta)}$$

By the theory of residues the free modes are

$$I(\varphi) = \frac{4\pi\ V\omega\epsilon (a^2+\bar{p}^2)^{1/2}}{v^2} \sum_{i=1,2} \left. \frac{e^{\,j\beta_i\varphi}}{\frac{d}{d\beta} F(\beta)} \right|_{\beta=\beta_i}$$

where $V = E\,\varphi_o \sqrt{a^2+\bar{p}^2}$ and use of the fact that

$$\frac{\sin \beta_i \frac{\varphi_o}{2}}{\beta_i \frac{\varphi_o}{2}} e^{j \beta_i \frac{\delta}{2\bar{p}}} \approx 1 \qquad \text{for } i = 1,2$$

The current on the helix is then written for $\varphi > 0$ as

$$I(\varphi) = I_o \left[e^{-j \beta_1 \varphi} + a\, e^{-j \beta_2 \varphi} \right]$$

A plot of $|a|$ for $\psi = 12.6$ and $\delta = .15$ as a function k' is shown in Figure 8.1.

8.2.2.2 The Second Problem

A finite helix $|\varphi| \leq \varphi_o$ is fed at $\varphi = 0$. The input impedance is given by

$$Z_{in} = \frac{< I, \quad I >}{I^2(o)}$$

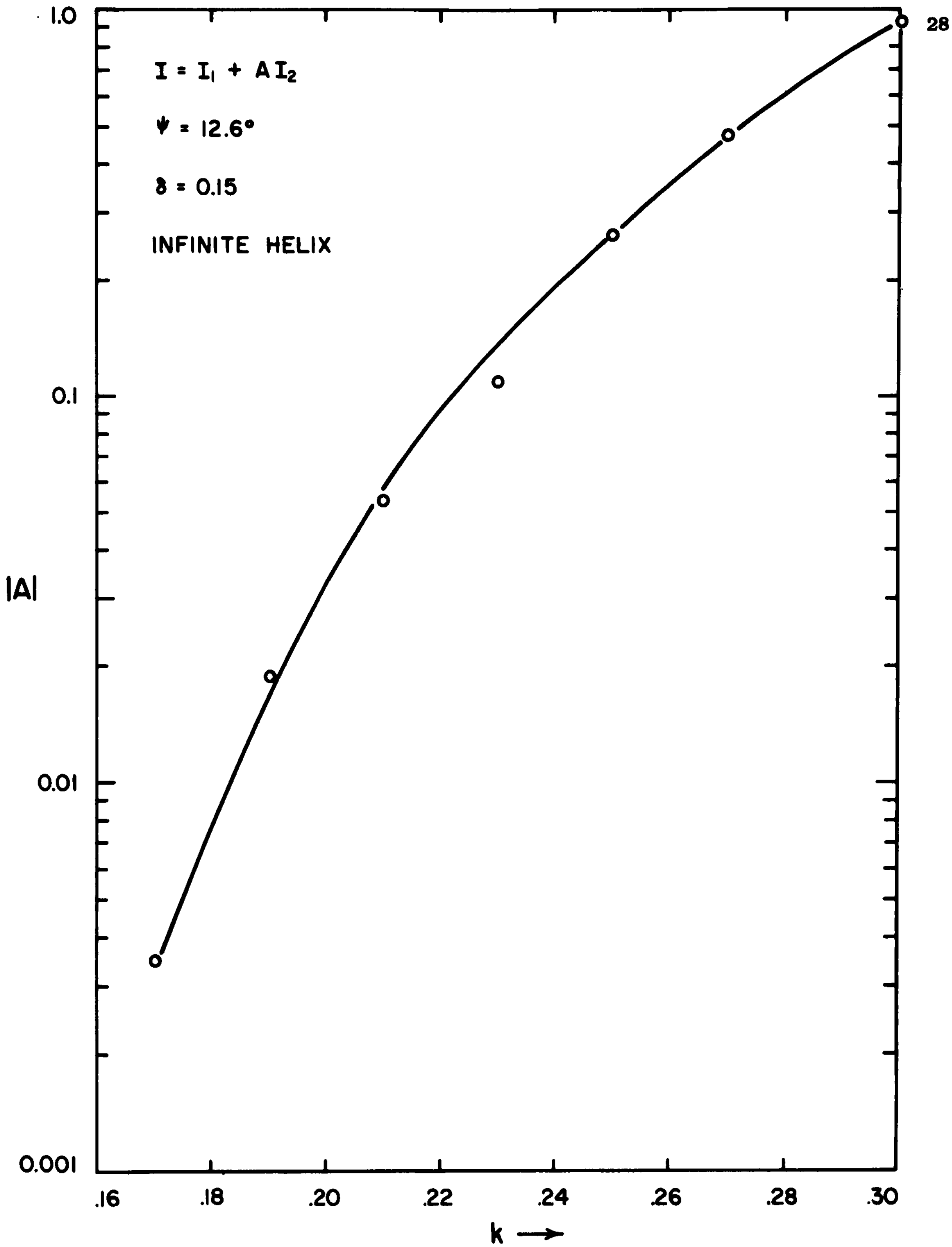
where

$$< A,B > = \int_{-\varphi_o}^{\varphi_o} AB d\varphi$$

$\mathbb{K}$ I is the integral operation on I to give $E_{\parallel}$ namely

$$\mathbb{K} I = \int_{-\varphi_o}^{\varphi_o} I(\varphi') \; K(\varphi-\varphi') \; d\varphi'$$

$$K(\varphi-\varphi') = \frac{1}{4\pi\omega\epsilon j (a^2+\bar{p}^2)^{1/2}} \left\{ \frac{\partial^2}{\partial\varphi^2} + k^2 [a^2 \cos(\varphi-\varphi') + \bar{p}^2] \right\} \frac{e^{jkR}}{R}$$



I = I₁ + A I₂
ψ = 12.6°
δ = 0.15
INFINITE HELIX
28
1.0
0.1
0.01
0.001
|A|
.16
.18
.20
.22
.24
.26
.28
.30
k

Figure 8.1

The equation for input impedance is stationary with respect to the current. This result has been demonstrated by Storer[1].

The current assumed is

$$I(\varphi) = I_1(\varphi) + A\, I_2(\varphi)$$

where I_1 and I_2 both satisfy the boundary conditions $I_1(\pm\, \varphi_o) = 0$ and correspond to the first and second mode respectively.

$$I_1 = \sin\ \beta_1(\varphi_o - \varphi)$$

$$I_2 = \sin\ \beta_2(\varphi_o - \varphi)$$

To find the conplex constant A one sets $\partial Z_{in}/\partial A = 0$. This then permits a solution for A

$$A = \frac{I_2(o)\mathcal{L}_{11} - I_1(o)\mathcal{L}_{12}}{I_1(o)\mathcal{L}_{22} - I_2(o)\mathcal{L}_{12}}$$

where

$$\mathcal{L}_{oj} = \langle I_i, \mathbb{K}\, I_j \rangle$$

These integrals were evaluated for a helix with $\psi = 12.6$ and $\delta = .15$ for $\varphi_o = 4(2\pi)$, $6(2\pi)$ and $8(2\pi)$. The results for $|A|$ are shown in Figure 8.2 as a function of frequency.

The periodic variations of $|A|$ with frequency is due to the variation of the input impedance with frequency.

Note that as the frequency is increased that the second mode increases as was calculated from the infinite case. This also agrees with experimental observations by Marsh.

8.3 Plans for the Next Interval

A technical report will be prepared.

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[1] J. E. Storer, "Variational Solution to the Problem of the Symmetrical Cylindrical Antenna," T.R. 101, Cruft Lab., Harvard University, 1950.

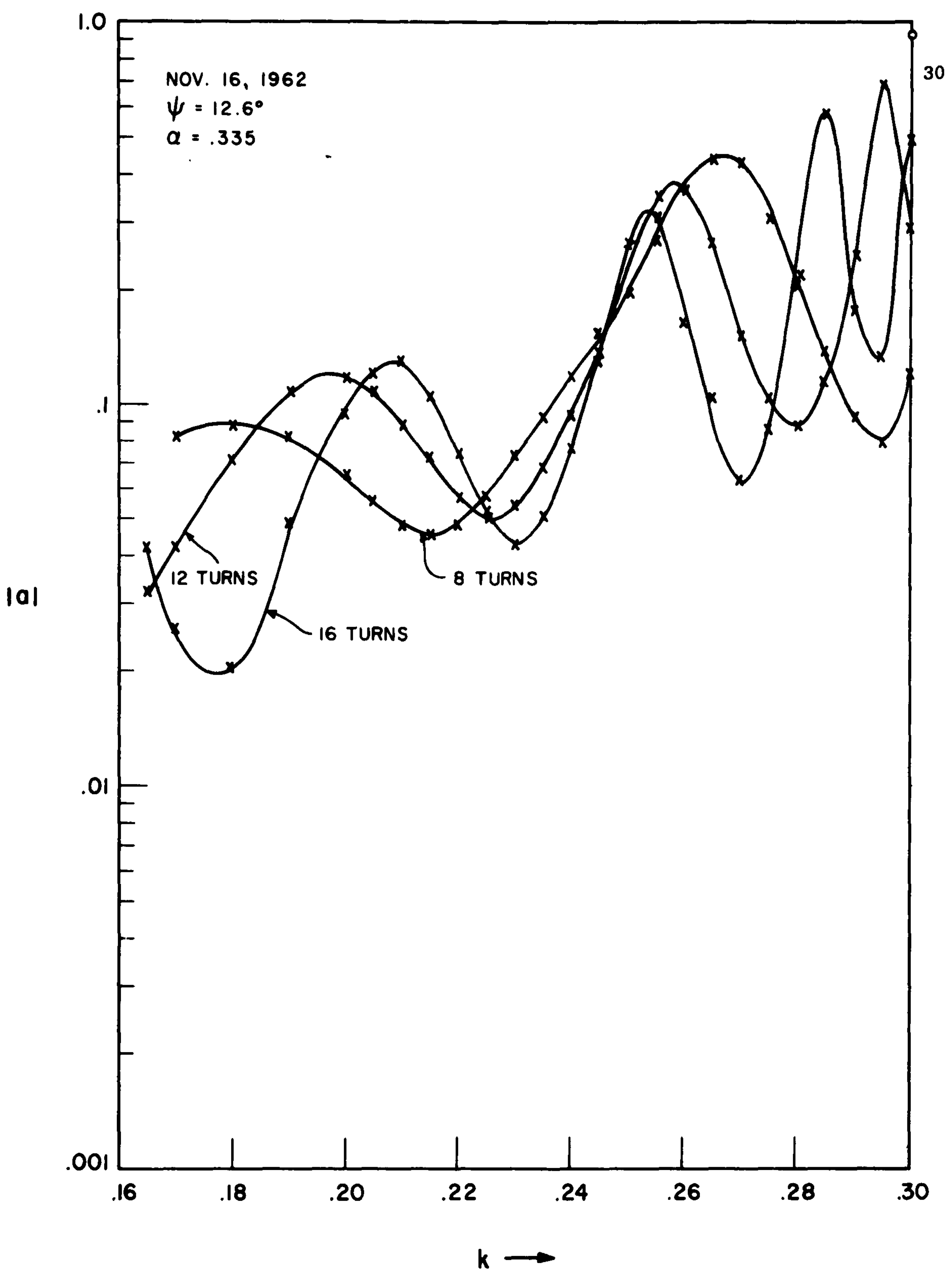
NOV. 16, 1962
ψ = 12.6°
α = .335
30
1.0
.1
.01
.001
|a|
12 TURNS
8 TURNS
16 TURNS
.16
.18
.20
.22
.24
.26
.28
.30
k

Figure 8.2

9. INVESTIGATION OF A CLASS OF PERIODIC STRUCTURES

9.1 Purpose

The purpose of this project is to study a class of periodic structures for wideband antenna applications.

9.2 Factual Data

9.2.1 Research Staff - S. Laxpati, M. Wahl, R. Mittra

9.2.2 Status

The different phases of the work carried out under the general heading of periodic structure study will be reported in four sections as follows.

(a) Study of uniform and log-periodically loaded transmission line

This investigation is a continuation of the work reported in Technical Report No. 59, Contract No. AF33(657)-8460, titled, "Theoretical Study of a Class of Log-Periodic Circuits."

The analysis of the periodically loaded transmission line was extended to the situation where the mutual impedances are also included (see Figure 9.1).

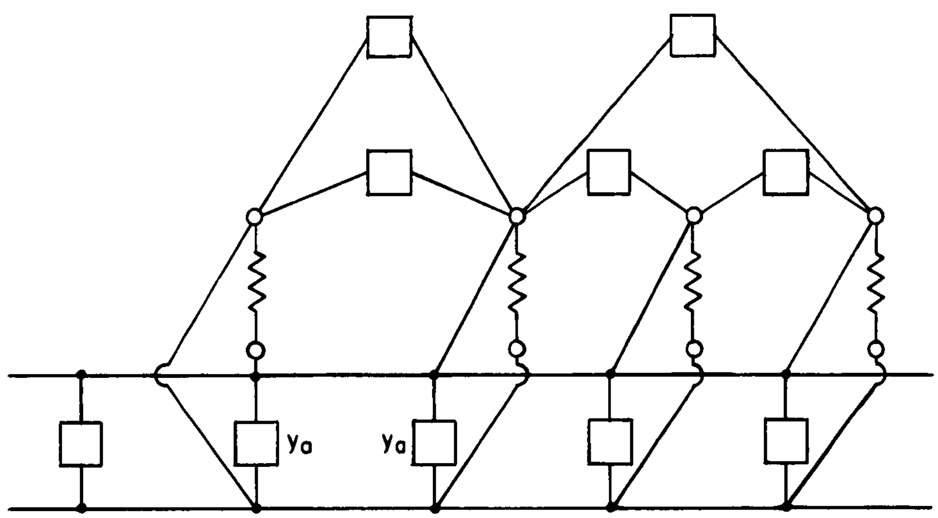

Figure 9.1

The characteristic equation for the propagation constant β is obtained by using an equivalent circuit of the type shown in Figure 9.2 which simplifies the problem considerably.

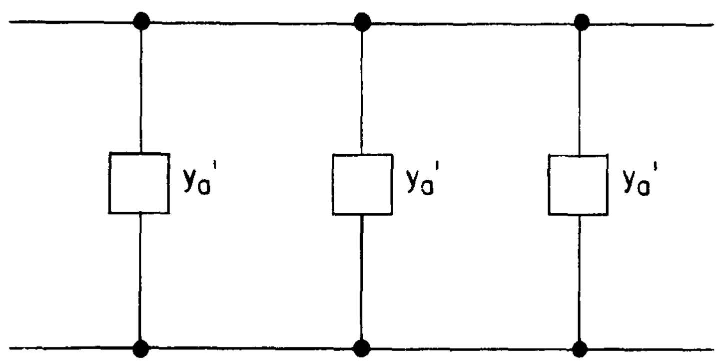

Figure 9.2

This new shunt admittance involves the mutual terms and the propagation constant β. The characteristic equation reads

$$\cosh \beta L = \frac{-j \sin kL}{2(Z_{11} + 2Z_{12} \cosh 2\beta L) + j \cos kL}$$

$k = 2\pi/\lambda$, β = propagation constant and Z_{11}, Z_{12} etc. are the impedance parameters of the loading circuit.

Theoretical calculations of k-β diagram for dipole loaded transmission lines have been performed and the results are now being compared with experiments reported by Mayes and Ingerson[2].

The analysis has also been extended to the log periodic case. The characteristic equation for the LP case reads

---

[2] P. E. Mayes and P. G. Ingerson, "Near-Field Measurements on Backfire Periodic Dipole Arrays," presented at the Fall URSL meeting, Ottawa, Canada, October, 1962. Also Proceedings of the 12th Annual Symposium on USAF Antenna Research and Development Program, Monticello, Illinois, October 1962.

$$\gamma(k\alpha^{n}) + \gamma(k\alpha^{n}) = 2\cos(kL\alpha^{n}) + j\,\frac{Y_{n} + Y_{n+1}}{2}\sin(kL\alpha^{n})$$

where $\gamma(k\alpha^{n})$ is $V_{pM/Vp}$ or the propagation constant for the pth cell, $a = \tau^{-1}$, the log period and

$$Y_{p} = \frac{\Gamma(k\alpha^{p-1})}{[\text{-----} + Z_{p,p-1}\,\gamma(k\alpha^{p-2}) + Z_{pp}\,\gamma(k\alpha^{p-1}) + Z_{p,p+1}\,\gamma(k\alpha^{p}) + \text{-----}]}$$

The solution of this equation will be undertaken in the future months.

(b) Investigation of uniform and tapered corrugated surfaces

The purpose of this part of the project is to apply the Brillouin diagram (k-β) for the design of a class of antennas.

The k-β diagram of a corrugated surface structure (Figure 9.3) was obtained experimentally and was compared with the theoretical curve computed earlier. It was predicted on the basis of the theoretical and experimental studies on this structure that, with the usual horn or loop feed at the input, the structure does not lend itself to a suitable broadband antenna design even when tapered in a log-periodic manner.

A new feed system was therefore devised which modifies the k-β characteristics of the structure to a desirable form. Several LP structures were designed using the new feed system. These antennas exhibited backfire radiation over a frequency range and appear promising as broadband radiators. Certain parameter studies are now being carried out to obtain better design information.

The tapered corrugated structure has been called a letter-rack antenna (Figure 9.4) and is a very suitable antenna for flush mounting on aircraft and also for burrying it flush with ground.

The future program is to continue the study of the k-β properties of the letter-rack and to obtain better design information on the antenna.

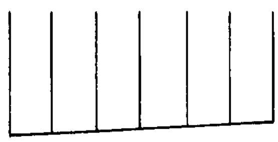

Figure 9.3 Corrugated surface structure

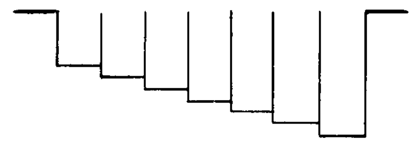

Figure 9.4 Letter-Rack Antenna-Tapered Corrugated Structure

(c) Modulated Periodic Structures

In this section we report the results of a study of a periodic wave-guide structure as a preamble to considering a more general type of modulated structure. The periodically loaded waveguide under consideration is shown in Figure 9.5.

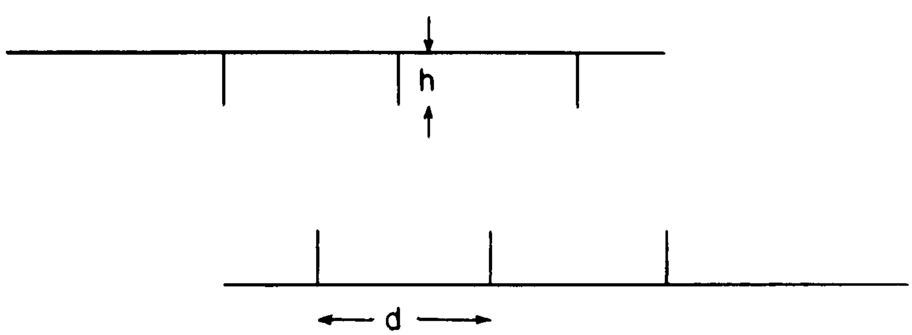

Figure 9.5

The interlacing of the loading diaphragms may be considered as a certain type of modulation.

Floquet's theorem which applies to the properties of periodic structures defines the period of such structures on the basis of translation alone and according to this the period of the structure should be 'd'. However, a generalization of this theorem is necessary when considering geometries of the type shown because of the presence of a glide reflection type of symmetry. An application of the generalized Floquet's theorem* obtains an effective period of $d/2$ i.e., if β is a solution of the propagation equation, then $\beta + 4n\pi/d$ $n = \pm 1, \pm 2, \pm 3 \ldots$ etc. are also solutions but not $\beta + 2p\pi/d$ for p odd. The field problem of the periodic structure under consideration was formulated exactly in terms of an infinite set of equations. The determinantal equation for β was studied and it was shown that $\beta + 2p\pi/d$ for p odd is not a solution which is in agreement with the prediction based on an application of the generalized Floquet's theorem. This theorem should also find application to the case of other periodic structures such as the dipole array.

Further investigation with the modulated periodic structures is planned for the future.

(d) Diffraction by a grating

The problem of field solution for a step discontinuity in a rectangular waveguide (Figure 9.6) has been the subject of investigation of a large number of workers. A solution to this problem is also applicable to the study of diffraction by a thick grating (Figure 9.7) and in addition is of interest to one engaged in the study of propagation along periodic structures (Figure 9.8).

The problem of the step discontinuity in a waveguide has been formulated in terms of a doubly infinite set of equations. A series solution of this set of equations has been obtained for the case of the ratio b/d as an integer.

The future program is to extend the solution to the rectangular bar grating problem and to other possible structures.

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*The generalized Floquet's Theorem is due to Professor G. A. Deschamps (unpublished work).

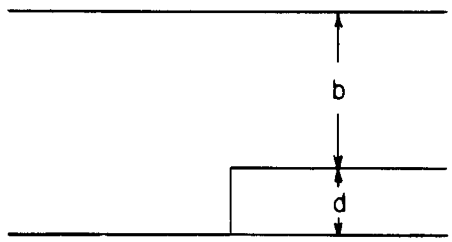

Figure 9.6 Step discontinuity in a waveguide

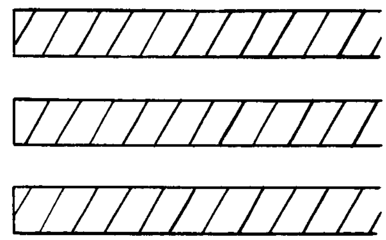

Figure 9.7 Thick grating

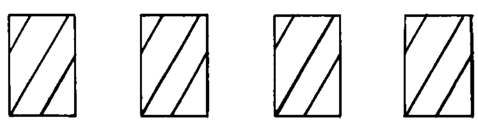

Figure 9.8a Grating of rectangular bars

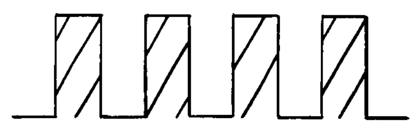

Figure 9.8b Thick corrugated structures

10.2.3 Papers and Publications

R. Mittra, "Relative Convergence of the Solution of a Doubly Infinite Set of Equations," Journal of Res. NBS (Accepted for publication).

K. E. Jones and R. Mittra, "A Study of Continuous and Log-Periodically Loaded Transmission Lines," presented at the Fall URSI Meeting, Ottawa, Canada, October, 1962.

R. Mittra and M. Wahl, "The Letter-Rack Antenna - a Flush Mounted Wide Band Antenna of Log-Periodic Design," accepted for presentation at the IRE National Convention, 1963.

R. Mittra and K. E. Jones, "Theoretical Brillouin (k-β) Diagrams for Monopole and Dipole Arrays and their Applications to Log-Periodic Antennas," accepted for presentation at the IRE National Convention, 1963.

10. FOCUSING OF RADIATION FROM A BUNCHED CERENKOV BEAM

10.1 Purpose

The purpose of this study is to investigate methods of coupling energy from fast wave line sources, such as Cerenkov radiators, directly into an antenna beam, a radiating aperture, or other device.

10.2 Factual Data

10.2.1 Research Staff - H. A. Shubert

10.2.2 Status

The design for a device to couple from a Cerenkov beam has been developed in this investigation and was described in the previous quarterly report[3]. It consists of a grating formed by cutting a slot in the form of an Archimedian spiral into a plastic sheet. The grating constant is adjusted so as to diffract the conical wave radiated by the Cerenkov beam in the direction normal to the sheet. By making the grating in the form of a spiral, the grating at the bottom is displaced from the grating at the top by one-half period, shifting the phase of the radiated wave by one hundred and eighty degrees, so that the waves from the top and bottom add in phase. The same applies to wave radiated by opposite sides of the grating. The result is a structure which produces a lobe normal to the sheet which is circularly polarized.

Such a grating has been constructed to work at a frequency of 16.0 Gc. It was formed by milling a spiral groove into a 1 in. thick sheet of plexiglass. The outer dimension of the grating was 24 in., or about 36λ (free space wavelengths). The depth of the slot was 5/8 in. as calculated by the formula $L = \lambda/2\ (n-1)$, where $n = 1.6$ is the refractive index of the plexiglass. The slot width and spacing were each 5/8 in. making the grating constant $2w = 1\ 1/4$ in. calculated from $2w = \lambda/\sin\theta$, where θ is the angle of radiation of the conical wave to be focused. Since fast wave line sources formed by slots cut in waveguides were used instead of a Cerenkov beam, and since the angle of radiation from such a source is given by $\sin\theta = \lambda/2a$, it turns out that $2w = 2a$, where a is the broad wall inner dimension of the waveguide. Thus, for the dominant mode exciting a given waveguide, the slot width and

[3]Quarterly Report No. 4, Contract AF33(657)-8460, On Research Studies Related to Antennas, Section 11, p. 19, 1 October 1962.

spacing are independent of frequency. The grating with the waveguide feed did indeed have patterns which were nearly frequency independent.

Experimental patterns for a spiral dielectric grating fed by an off-center longitudinal waveguide slot in the broad wall produced a double lobe normal to the grating. One of the lobes was precisely along the normal, while the other was slightly off on the side of the slot, and about 4 db down from the main lobe. This effect has been ascribed to non-uniform illumination of the grating, with stronger illumination near the edges, thus causing an interferometer effect. The non-uniform illumination seems to be due primarily to interaction of the grating with the slot, rather than the exponential decay due to energy leakage along the guide, as neither a very thin slot nor a tapered slot seemed to reduce the side lobe very much.

The interaction of the grating with the slot seems to be important only when the polarization of the incident conical wave E field is parallel to the slots, as is the case with the longitudinal slot radiator in the broad wall of the waveguide. With a uniform array of transverse slots cut in the narrow wall of the waveguide, the E field is polarized parallel to the generators of the conical wave. The pattern of the grating fed by a uniform array of slots cut in the narrow wall of the guide is shown in Figures 10.1 and 10.2. Fifty slots were cut across the narrow wall and angled at 5° so they would radiate. The slots were 1/16 in. wide and spaced 1/4 in. apart. The pattern cut is taken in this horizontal plane, with horizontal polarization. That there is still some interaction between the waveguide feed and the grating is evidenced by the two symmetrical side lobes 8 db down, indicating a slight interferometer effect. In the other polarizations, or other cuts, which are not shown, the pattern has no side lobes above 25 db except those near the angle of the original conical wave, and the back lobe. Only with this cut and polarization is the radiation from both sides of the lens effective in forming the pattern, so that only in this pattern is the interferometer effect due to the edges visible.

10.3 Plans for the Next Interval

The spiral dielectric grating was the most practical of several devices considered as couplers from a fast wave line source. Having completed experimental testing on the spiral dielectric grating, work on this project is to be discontinued.

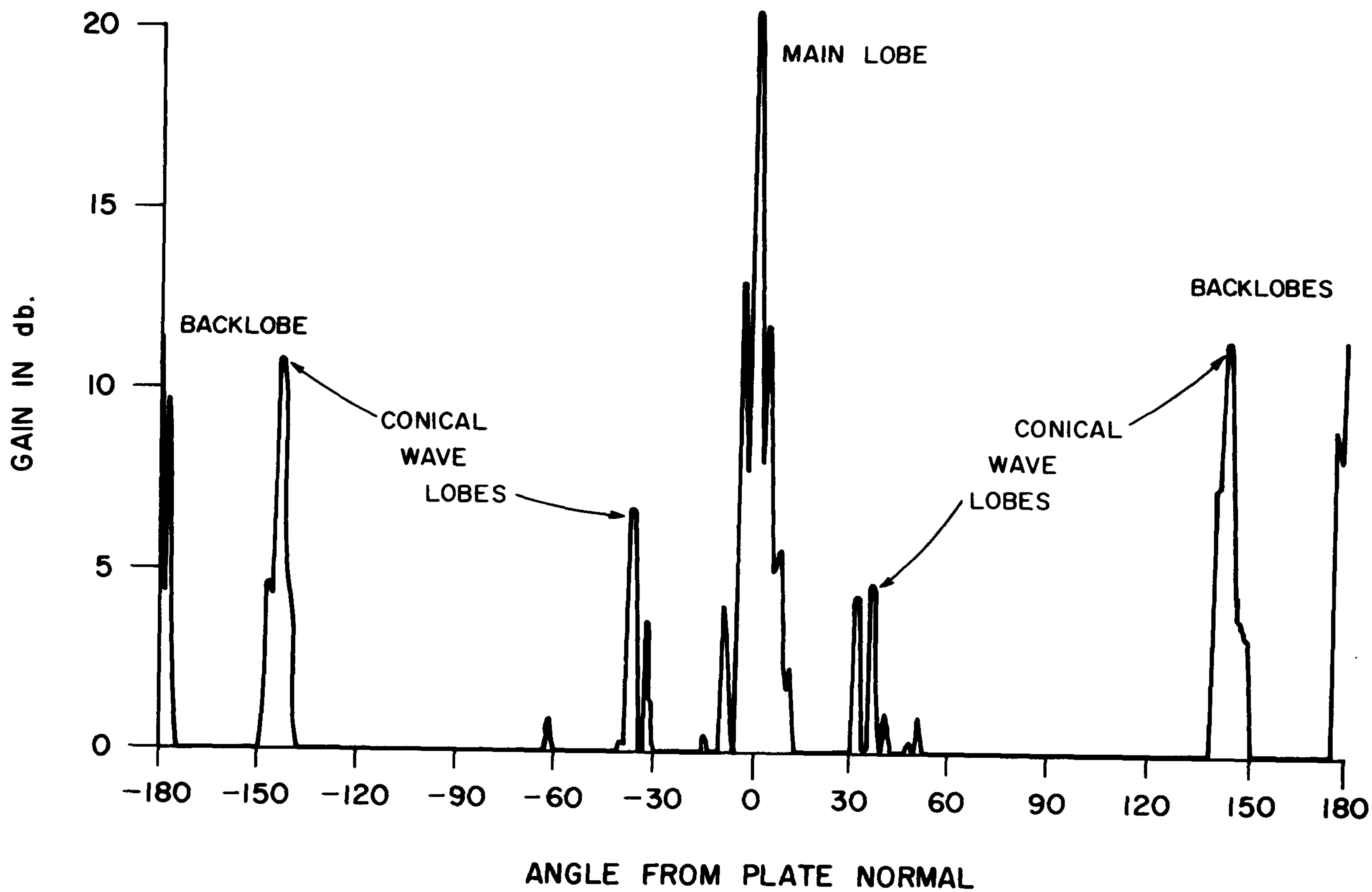

Figure 10.1. The Pattern of the Archimedes Plastic Spiral Fed By a Uniform Array of Slots

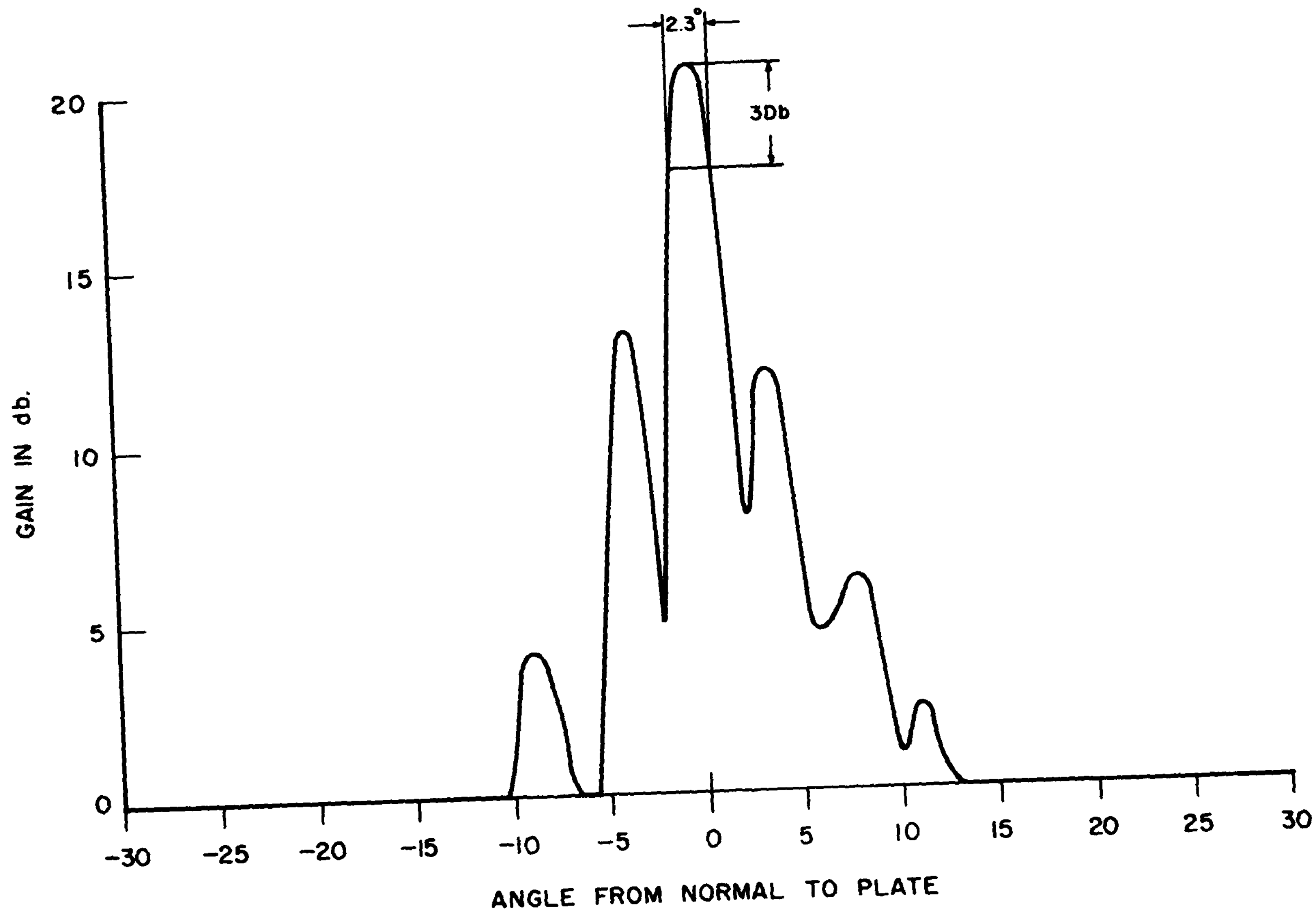

Figure 10.2. Detail of the Main Lobe of the Pattern Shown in Figure 10.1.

11. TRANSMISSION BETWEEN ANTENNAS WHEN THE FAR FIELD APPROXIMATION DOES NOT APPLY

11.1 Purpose

The purpose of this research is to determine the transmission coefficient between antennas located in their respective near fields.

11.2 Factual Data

11.2.1 Research Staff - J. R. Pace, G. A. Deschamps

11.2.2 Status

In the last quarterly report, a device was suggested which would measure the far field patterns of an antenna when it was situated in the antenna's near field. The device was to be a specially designed end-fire array of elements. Mention was made that further study would be needed to determine the workability of the device. It was decided that the array has a fundamental drawback. First, it was designed on the basis that there was no interaction between the elements of the array itself. That is, the spacing between elements was relatively large. The drawback of course, is that as the array is made longer in terms of wavelengths, its own gain increases. This tends to defeat the purpose of the array. Yet, if the interelement spacing is small, the elements have considerable interaction. Computations which would show that the device theoretically worked have not yielded positive results. For the present, work on this device is discontinued.

Attention is now directed to the transmission coefficient, and its applicability to calculation of the coupling between specific apertures located in their respective Fresnel regions. In the past few years, a number of authors have developed expressions for the so-called "near-field gain" of apertures. Their results are limited in the sense that they generally involve complex integrals which are difficult to evaluate. The advantage of the present approach is that the problem is reduced to that of evaluating the Fourier transform which yields the far field. This integral is usually easy to evaluate compared to integral which represents the Fresnel field.

A computer program has been written which gives the transmission between large square apertures. The results will show the effects of variations in the relative angular orientation of the apertures, as well as the effects of variations in the spacing between elements. The scalar form for the transmission coefficient has been used. Linear polarization has been assumed. Fresnel region coupling is to be evaluated.

11.3 Plans for the Next Interval

The results from the computer will be reduced to graphical form. Where possible, these results will be compared with those of other authors.

12. RADIATING LENS ILLUMINATED FROM A GOUBAU BEAM-WAVEGUIDE

12.1 Purpose

The purpose of this project is to determine the feasibility of developing a high resolution millimeter antenna by coupling from a Goubau beam-waveguide to a radiating lens.

12.2 Factual Data

12.2.1 Research Staff - M. Fournier, P. E. Mast

12.2.2 Status

The design of a "lens" coupling system from a Goubau beam-waveguide to a radiating aperture based on a geometrical optics approach has been completed. The computer program for calculation of the lens surfaces has been run, and the required lens holders, etc., are in the final stages of construction.

The Goubau beam waveguide has been completely assembled, checked out, and is operating satisfactorily at 70 Gc. The indoor pattern range has been put in operation at 70 Gc.

Theoretical work on the losses in the "lens" and the side lobes to be expected from the "lens" coupling system has been continued.

12.3 Plans for the Next Interval

In the next interval mechanical construction of the lens coupler will be completed and its radiation patterns measured. Theoretical work on the effects of losses, side lobe level, etc. will be continued.

13. ZONE PHASE PLATES

13.1 Purpose

The purpose of this project is to investigate the properties and uses of zone phase plates.

13.2 Factual Data

13.2.1 Research Staff - H. A. Shubert

13.2.2 Status

The problem under consideration is the calculation of the diffraction of a plane wave by a zone phase plate composed of an infinite periodic grating of bars of rectangular cross section. The bars have zero conductivity, relative dielectric constant ϵ_2, and relative permeability μ_2. They have a thickness 2t and are spaced 2s, giving a grating constant $2w = 2(s+t)$. The depth of the bars, through which an incident wave must pass, is L. By making the bars infinitely long in the y direction, and requiring that the plane wave be incident normal to the y direction, the problem is made two dimensional. Field components E_y, H_x, and H_z become uncoupled from field components H_y, E_x, and E_z thereby allowing each set of three components to be treated separately. All results for the second set may be derived from results for the first set by duality considerations.

The problem was formulated for general direction of incidence (in the x,z plane) of the plane wave, general bar thickness t and spacing s, and general bar width L. It was possible to separate the diffracted fields into an infinite number of discrete self-consistent modes $\phi_m(x)\ e^{\pm j\sigma_m z}$ in front of and in back of the plate, and another infinite discrete set of self-consistent modes $\psi_p(x)\ e^{\pm j\tau_p z}$ in the plate. By imposing the boundary conditions upon tangential fields at the front of the plate, $z = 0$, and the back, $z = L$, it was possible to formulate the problem either as an infinite set of simultaneous equations for the mode coefficients, or as an integral equation for the tangential fields at the front and back of the plate. Either formulation may be formally solved by well-known inversion methods. In both cases, the practical calculation of the fields is hampered by the complicated nature of the matrix coefficients or the kernel functions. The convergence properties of either are generally unknown.

By making the assumption, which is generally true for a zone phase plate, that the period of the grating 2w corresponds to two of the Fresnel zones produced on the front surface of the plate by the incident plane wave, the problem is simplified. All fields may then be classified as either even or odd about the center of any one of the bars in the x direction. The even and odd fields do not couple, and therefore may be treated in separate problems. Odd fields do not contribute to a main lobe off the normal to the plate, and produce only side lobes.

By making the further assumption that s = t, as is also true in most zone phase plates, the asymptotic form of the elements in the formal solution

$$R_{mp} = \int_{-w}^{w} \frac{\phi_m(x)\ \psi_p^*(x)}{\mu(x)}\, dx$$

$$S_{mp} = \int_{-w}^{w} \phi_m(x)\ \psi_p^*(x)\, dx$$

for large p is very much simplified. For higher order cutoff modes in the plate, of which only those involving odd functions will be reproduced, the asymptotic form for large p is:

$$R^o_{mp} = \begin{cases} \frac{w}{2}\left(1 + \frac{\mu_o}{\mu_2}\right) \delta^p_m, & p_{odd},\ m_{odd} \\ \frac{2w}{\pi}\left(1 - \frac{\mu_o}{\mu_2}\right)(-)^{m+p-1/2}\ \frac{m}{m^2 - p^2}, & p_{odd},\ m_{even} \\ w\ \delta^p_m & p_{even} \end{cases}$$

$$S^o_{mp} = \begin{cases} w\ \delta^p_m, & p_{odd} \\ -\frac{2w}{\pi}\left(1 - \frac{\mu_2}{\mu_o}\right)(-)^{m+p-1/2}\ \frac{p}{m^2 - p^2}, & p_{even},\ m_{odd} \\ \frac{w}{2}\left(1 + \frac{\mu_2}{\mu_o}\right) \delta^p_m, & p_{even},\ m_{even} \end{cases}$$

The above expressions, relating to TE fields, may be changed to expressions for TM bields by substituting ϵ's for μ's. In the TE case, with $\mu_2 = \mu_o$ and p large

$$R_{mp} \longrightarrow S_{mp} \longrightarrow w\ \delta_m^p$$

so that the matrices and/or series involved in the formal solutions can be truncated. At least for this last case, the solutions for the scattered fields are not difficult.

So far, no account has been taken of the effect of bar width L. Preliminary experimental results do not seem to show the large effect from this parameter. Also, experimental results on zone plates in which alternate zones are blocked with aluminum foil seem to indicate better performance, with regard to side lobe level, than is indicated by physical optics considerations.

13.3 Plans for the Next Interval

Work will continue on this project, and will aim to calculate patterns. Experimental work will be done and compared to calculations.

14. STUDY OF FOCUSING PROPERTIES OF METAL-LENS ANTENNAS

14.1 Purpose

The purpose of this investigation is to solve rigorously for the fields around the focal point of a plano-elliptic metal plate lens. In particular, the chromatic aberration of the lens is to be studied.

14.2 Factual Data

14.2.1 Research Staff - J. R. Pace, R. Mittra

14.2.2 Status

Since the pioneering work of W. E. Kock who developed the metal-lens antenna, considerable effort has been extended towards determining rigorously the properties of the metal plate medium. (The original lenses of Kock were designed employing geometrical optics). To date, successful studies have been made on a medium consisting of plates of semi-infinite extent. Reflection and transmission coefficients have been calculated for this medium, and an attempt has been made to apply these results to the practical lens. The true lens, of course, consists of plates of finite and unequal length. The problem of analyzing the true lens, it is believed, has not yet been solved.

This problem is presently being formulated. The solution of it will lead to a knowledge of the fields in the neighborhood of the focal point of the lens. Moreover, the chromatic aberration in the lens can be analyzed rigorously. A comparison with the results predicted by geometrical optics can also be made.

14.3 Plans for the Next Interval

Formulation of the problem will be completed and computation of the fields will be started. The use of the University of Illinois' digital computer is contemplated.

15. MILLIMETER ANTENNA LITERATURE

15.1 Purpose

The purpose of this survey was to determine the literature available on the subjects of millimeter wave transmission and radiation.

15.2 Factual Data

15.2.1 Research Staff - J. F. Kauffman

15.2.2 Status

In addition to the standard technical journals, the Engineering Index was surveyed for the years 1953 to 1962 for all articles on millimeter wave transmission and millimeter antennas. The ASTIA Technical Abstracts Bulletins for 1961 and the first three quarters of 1962 were searched for reports not published or referenced in the technical journals.

The general problems encountered at millimeter wavelengths in conventional transmission systems are the necessity of very tight tolerances for components, and high attenuation. Most of the research, therefore, has been directed toward developing new methods of transmission which circumvent these difficulties. Some of these schemes are:

a) Goubau beam-waveguide
b) Non-waveguide components
c) H-guide
d) "O" and "X" guide
e) Dielectric lines
f) Multimode transmission in circular waveguide

The Goubau beam-waveguide[1] is a free-space transmission system in which the phase of the wave is corrected periodically with dielectric lenses. The non-waveguide components[2] are devices which allow free space waves to be channeled in much the same manner as are guided waves in a waveguide circuit. One such component is a 90° bend made of a totally reflecting dielectric prism; another component is a 3 db-coupler made from a half-reflecting plane mirror. H-Guide[3,11] is constructed from two parallel metallic planes with a small dielectric strap running down the center of the guide. This guide has good loss properties and is easy to construct. The "O" and "X" guides[4] and the dielectric lines[6] are all in the same category with the "O" and "X" lines being merely special configurations of dielectric lines. These lines have the common disadvantage that the guided waves are loosely coupled and therefore tend to radiate at the slightest discontinuity in the line. The circular waveguide operated multimode[5,7] also has low losses. However, it has the disadvantage of requiring very pure mode feeding.

Less literature is available on millimeter wave antennas[14,15] than on millimeter wave transmission methods. The work which has been done has been oriented toward scaling antennas which were designed for lower frequencies. Because of this, the tight tolerance and high loss problems are still present.

A listing of all the reports and articles found in this survey is too lengthy to include here. The bibliography given below contains selected articles intended to give a general view of the types of work being done and the organizations doing this work.

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7. "A Note on Quasi-Optical Methods at Millimeter Wavelengths," L. Lewin, Proceedings of the Symposium on Millimeter Waves, April 1959, pp. 469-477, Brooklyn Polytechnic Institute.

8. "Millimeter Wave Research," Cotton, Sobel, Wedel, February 28, 1962, Electronic Communications, Inc., Contract AF30(602)2457, Project 4506.

9. "Research in Components and Techniques for Electromagnetic Warfare," Quarterly Report No. 9, 1 January to 31 March 1962, Radiation Laboratory, John Hopkins University, Contract AF33(616)6753, AD-326 273 (Secret)

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15. "Millimeter Wavelength Antenna Studies," Stanford Research Institute, Quarterly Report 8, 1 January to 1 April 1953, Contract DA 36-039-SC 5503.

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